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# Performance of Space-Time Coding with Hiperlan/2 and IEEE 802.11a WLAN Standards on Real Channels

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**Abstract** – In this paper we examine the performance of a space-time coded (STC) OFDM system. As an example, two emerging WLAN standards (IEEE 802.11a and ETSI Hiperlan/2) are modified to accommodate the STC technique. Performance results using statistical multi-element channel models are compared with those using state-of-the-art indoor MIMO channel measurements. We conclude that the gap between the ideal setting and our measured channel is 4, 5 and 6 dB for 1, 2 and 3 receive antennas respectively. We observe also that STC with (2Tx, 3Rx) over the measured channels doubles the capacity as compared to the equivalent mode of Hiperlan/2.

## I. INTRODUCTION

IEEE 802.11a and ETSI Hiperlan/2 represent two emerging wireless LAN standards that have been defined for the provision of high data rate services. Close co-operation between ETSI and IEEE has ensured that the physical layers of the two standards are harmonised to a large extent. Both standards are based on an OFDM modulation format, which utilises a cyclic prefix in order to render a wideband frequency selective channel into a number of parallel independent frequency non-selective channels.

Recent diversity developments have shown that gains similar to receive diversity are now possible using transmitter diversity [1]. The pioneering work in [1] ignited tremendous interest in Space-time trellis coded modulation (STTCM) schemes amongst the research community. This has led to a number of new STTCM schemes with improved performance [3,4,5]. The relative simplicity of combined OFDM and STTCM has been noted in [2] as a method of applying transmit diversity in wideband systems.

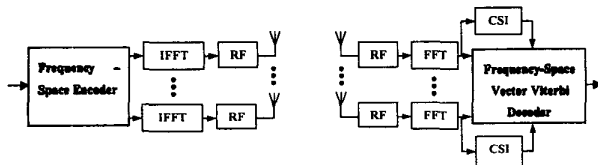


Fig. 1. STTCM applied to OFDM system (FSC-OFDM).

The aforementioned contributions investigated STTCM and STTCM-OFDM analytically and via means of simulations with synthetic channels (simple simulated channels that assume perfect decorrelation between elements). In this paper

we take the concept presented in [3] and apply it to measured wideband MIMO (Multiple Input Multiple Output) channels.

The presented results measure the performance of STTCM schemes when applied to OFDM systems where coding is performed in parallel across sub-carriers and antenna elements. These architectures are more aptly named Frequency-Space Coding (FSC). Figure 1 illustrates such an architecture. For low mobility, high data rate systems, the greater diversity available in the frequency domain compared to the time domain, can be more efficiently exploited by FSC-OFDM. Reference [8] provides a review of space-time coding as applied to broadband wireless systems.

This paper is organised as follows: section II briefly presents the measurement set-up and indoor environment. Section III develops the signal model for FSC-OFDM. Section IV reviews some known design criteria for STTCM and comments on applications to OFDM. Finally section V presents numerical results for the performance of FSC-OFDM over measured multi antenna element radio channels.

## II. MIMO CHANNEL MEASUREMENTS SETUP

The wideband MIMO measurements utilised here have been taken using a customised Medav RUSK BRI vector channel sounder operating in the 5.2 GHz band with 120 MHz of bandwidth. Each complete MIMO (8 Tx by 8 Rx) snapshot of the channel takes 102.4  $\mu$ s, which is well within the coherence time of the channel [10].

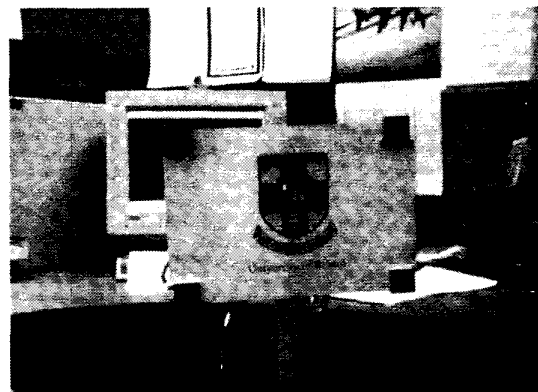


Fig. 2. "Laptop-sized" four double polarized element antenna array used for the MT.

The channel sounder uses a multitone signal comprising of 97 frequency points. A subset of 17 consecutive points is used to interpolate the 64 point FFT grid over a 20 MHz bandwidth as required for 802.11a and Hiperlan/2.

The measurements were taken in an open plan office with approximate dimensions of 30L x 20W x 4H (m). The Mobile Terminal (MT) antenna array comprises of 4 dual polarised patch antennas as depicted in Figure 2. The antennas were located in the corners of a dielectric plane with dimensions similar to that of a typical laptop computer cover. The Access Point (AP) utilises a circular array of printed dipoles – Figure 3. The AP antenna was located ~0.5m below the ceiling in the middle of the office.

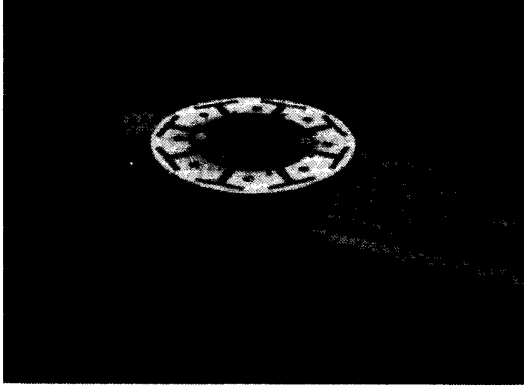


Fig. 3. Circular array of printed dipoles used for the AP.

### III. SYSTEM MODEL OF THE FREQUENCY-SPACE CODED OFDM

An OFDM system with a Cyclic Prefix (CP-OFDM) is assumed here, where the  $i^{\text{th}}$  transmitted block of data,  $\bar{\mathbf{u}}_i$ , is given by  $\bar{\mathbf{u}}_i = \mathbf{T}_{cp} \mathbf{F}^{-1} \mathbf{u}_i$ . The data vector,  $\mathbf{u}_i$ , is of length  $K$ , the size of the CP insertion matrix,  $\mathbf{T}_{cp}$ , is  $P \times K$ , where  $P = C + K$ ,  $C$  represents the length of the CP, and the Fourier transform matrix  $\mathbf{F}$  is of size  $K \times K$ . The receiver receives the current transmitted block of data  $\bar{\mathbf{u}}_i$ , in addition to a fraction of the previous block  $\bar{\mathbf{u}}_{i-1}$  through the excess length of the channel impulse response. This is described by Toeplitz channel matrices  $\mathbf{H}_0$  and  $\mathbf{H}_1$ , and the received signal block pertaining to  $\mathbf{u}_i$  is given by:

$$\bar{\mathbf{x}}_i = \mathbf{H}_0 \bar{\mathbf{u}}_i + \mathbf{H}_1 \bar{\mathbf{u}}_{i-1} + \bar{\boldsymbol{\eta}}_i \quad (1)$$

Both the above channel matrices are of size  $P \times P$  and are given by:  $(h_0, \dots, h_{L-1}, 0, \dots, 0)^T$  for the first column and  $(h_0, 0, \dots, 0)$  for the first row of  $\mathbf{H}_0$ ;  $(0, \dots, 0)^T$  for the first column and  $(0, \dots, h_{L-1}, \dots, h_1)$  for the first row of  $\mathbf{H}_1$ . Additionally, it is assumed that the length of the CP is  $C \geq L-1$ . The receiver removes the first  $C$  entries of  $\bar{\mathbf{x}}_i$  that are affected by  $\bar{\mathbf{u}}_{i-1}$  (Inter-Block-Interference - IBI). This is achieved by pre-multiplication with a matrix  $\mathbf{T}_R$  defined as:  $\mathbf{T}_R = [\mathbf{0}_{K \times C}, \mathbf{I}_{K \times K}]$ . Hence the input-output relationship can be expressed as:

$$\mathbf{x}_i = \mathbf{F} \mathbf{T}_R \mathbf{H}_0 \mathbf{T}_{cp} \mathbf{F}^{-1} \mathbf{u}_i + \mathbf{F} \boldsymbol{\eta}_i \quad (2)$$

where  $\boldsymbol{\eta}_i$  represents the ubiquitous additive noise vector. A specific construction of  $\mathbf{T}_{cp}$  guaranties that the concatenation  $\mathbf{T}_R \mathbf{H}_0 \mathbf{T}_{cp}$  is circulant, and thus is diagonalised by  $\mathbf{F}$ . Hence:  $\mathbf{F} \mathbf{T}_R \mathbf{H}_0 \mathbf{T}_{cp} \mathbf{F}^{-1} = \tilde{\mathbf{H}} = \text{diag}\{h(\omega^0), \dots, h(\omega^{K-1})\}$  and now:

$$\mathbf{x} = \tilde{\mathbf{H}} \mathbf{u} + \tilde{\boldsymbol{\eta}} \quad (3)$$

where  $\tilde{\boldsymbol{\eta}} = \mathbf{F} \boldsymbol{\eta}$ . We have dropped the block index “ $i$ ” since we are no longer concerned with IBI.

Next we follow the representation of the STTCM codes and the encoding process proposed in [3]. This is now extended to the case of STC-OFDM. We assume  $M$ -ary PSK constellations with a bandwidth efficiency  $m = \log_2(M)$  bits/sec/Hz. Consider a stream of binary data,  $\mathbf{d} = (d_1, d_2, \dots, d_n)$ , to be transmitted. The ST encoder will be defined by a generating matrix  $\mathbf{G}$  with  $n_T$  columns and  $m + s$  rows (the design criteria for  $\mathbf{G}$  and examples are provided in section IV), where  $s$  influences the memory (number of states) of the encoder. The entries of  $\mathbf{G}$  lie in the range 0 to  $M-1$ . Let  $\mathcal{M}$  denote

the mapping operation, e.g. for  $M$ -ary PSK:  $\mathcal{M}(x) = \exp(j2\pi x/M)$ . The frequency-space codeword  $\mathbf{c}$  at the carrier  $k$  is obtained from:

$$\mathbf{c}_k = \mathcal{M}(\mathbf{d}_k \mathbf{G} \pmod{M}) \quad (4)$$

Where:  $\mathbf{d}_k = (d_{mk+(m-1)} \dots d_{mk} \dots d_{mk-s})$  denotes the  $m + s$  long stream of input bits. In order to facilitate the description of an FSC-OFDM system with  $n_T$  transmit and  $n_R$  received antennas, a  $K \times n_T \times 1$  dimensional vector is defined:

$$\boldsymbol{\mu} = \text{vec}(\mathbf{c}) \quad (5)$$

Where:  $\text{vec}(\cdot)$  denotes the vector stacking operator. Additionally, a  $K \times n_R \times 1$  dimensional vector is defined as a stacked vector of the received data  $\boldsymbol{\chi} = [\mathbf{x}^{1,T} \dots \mathbf{x}^{n_T,T}]^T$ , and  $\boldsymbol{\eta}'$  is defined in a similar fashion. Recalling the result in (3), the FSC-OFDM model can be defined as:

$$\boldsymbol{\chi} = \mathcal{H} \boldsymbol{\mu} + \boldsymbol{\eta}' \quad (6)$$

With the overall channel matrix given by:

$$\mathcal{H} = \begin{bmatrix} \tilde{\mathbf{H}}_{1,1} & \tilde{\mathbf{H}}_{1,2} & \dots & \tilde{\mathbf{H}}_{1,n_T} \\ \tilde{\mathbf{H}}_{2,1} & \tilde{\mathbf{H}}_{2,2} & \dots & \tilde{\mathbf{H}}_{2,n_T} \\ \vdots & \vdots & \ddots & \vdots \\ \tilde{\mathbf{H}}_{n_R,1} & \tilde{\mathbf{H}}_{n_R,2} & \dots & \tilde{\mathbf{H}}_{n_R,n_T} \end{bmatrix} \quad (7)$$

Where: diagonal sub-matrices  $\tilde{\mathbf{H}}_{m,n}$  represent the frequency response of a channel between the  $n^{\text{th}}$  transmit and the  $m^{\text{th}}$  receive antenna.

In an AWGN channel and for PSK constellations, the maximum likelihood decoder can be realised using the Viterbi algorithm using the modified Euclidean distance metric:

$$\sum_{k=0}^{K-1} \left\| \mathbf{C}_k^{(M)^T} \mathbf{x} - \mathbf{C}_k^{(M)^T} \mathbf{H} \mathbf{C}_k^{(N)} \mathbf{e}_k \right\|^2 \quad (8)$$

The block diagonal selection matrices  $\mathbf{C}_k^{(M)}$  and  $\mathbf{C}_k^{(N)}$  simply select the appropriate received symbols and channel frequency response values for the metric computation for the  $k^{\text{th}}$  frequency index. Each matrix is constructed from vectors  $\mathbf{c}_k$  that have "1" at the  $k$  position and "0" elsewhere. Matrix  $\mathbf{C}_k^{(M)}$  has size  $n_R \times K$  blocks and  $\mathbf{C}_k^{(N)}$   $n_R \times K$  blocks.

In the above model we have assumed that the length of the codeword is equal to the number of sub-carriers  $K$ . Extension to longer codewords is straightforward. Also for the sake of clarity we have omitted the interleavers.

#### IV. REVIEW OF THE STTCM DESIGN CRITERIA FOR THE CASE OF OFDM

When STTCM is applied to OFDM systems the coding takes place across frequency and space rather than time and space – according to the model developed in section III. In the time domain the amount of available diversity is related to the Doppler phenomenon. Hence for low mobility high data rate systems (as considered here), the channel remains almost constant over a frame. Conversely, delay spread in the radio channel gives rise to diversity in the frequency domain. It is expected that WLAN systems will operate in environments ranging from frequency flat to frequency selective. Hence, the STTCM code designed for OFDM should perform well in quasi-static channels in addition to fast fading channels. A fast fading channel (in the frequency domain) will occur, for example, with ETSI specified channel model E [7], especially when an interleaver is adopted.

##### A. Minimum determinant criterion

The first space-time codes were designed using criteria developed for quasi-static Rayleigh fading channels [1]. The pairwise error probability for this case is minimised when the minimum rank of  $\mathbf{B}$  and the minimum product of the non-zero eigenvalues  $\lambda$  of the distance matrix  $\mathbf{A} = \mathbf{B}^H \mathbf{B}$  is maximised [1]. For full rank codes this amounts to maximising the minimum determinant of the distance matrix  $\mathbf{A}$ .

$$\mathbf{B}(\mathbf{c}, \mathbf{e}) = \begin{bmatrix} e_1^1 - c_1^1 & \cdots & e_1^{n_r} - c_1^{n_r} \\ \vdots & \ddots & \vdots \\ e_K^1 - c_K^1 & \cdots & e_K^{n_r} - c_K^{n_r} \end{bmatrix}$$

$$P(\mathbf{c} \rightarrow \mathbf{e}) \leq \left( \prod_{i=1}^R \lambda_i \right)^{-n_r} (E_s / 4N_0)^{-R n_r}$$

It has been shown that these codes continue to perform well in Rician channels [1]. Furthermore, it has been proved in [9] that these codes provide at least the same order of diversity over fast fading channels, as compared to slow fading channels.

This result is extremely useful for FSC-OFDM. Examples of codes constructed to maximise the *determinant criterion* include the codes of Tarokh, Seshadri and Calderbank [1] denoted as TSC, and the codes of Baro, Bauch, and Hansmann [3] denoted as BBH codes – Table 1.

##### B. Minimum product distance criterion

Reference [1] also develops the design criteria for fast fading channels. For this case, the diversity gain is maximised when the number of time instances in which two codewords differ is maximised. The coding advantage is governed by the minimum value of the product of the distances for which the codewords are different.

$$P(\mathbf{c} \rightarrow \mathbf{e}) \leq \prod_{k \in V(\mathbf{c}, \mathbf{e})} \left( \frac{|\mathbf{c}_k - \mathbf{e}_k|^2 E_s}{4N_0} \right)$$

This *minimum product distance* has to be maximised to achieve maximal coding advantage. The codes of Firmanto, Vucetic and Yuan denoted as FVY were specifically designed to optimise the performance over fast fading channels.

Table 1. Various 4PSK STTCM codes studied in this paper.

REF	NO. STATES	GENERATING MATRIX $\mathbf{G}^T$
[1] TSC	4	$\begin{bmatrix} 0 & 0 & 2 & 1 \\ 2 & 1 & 0 & 0 \end{bmatrix}$
[1] TSC	8	$\begin{bmatrix} 0 & 0 & 2 & 1 & 2 \\ 2 & 1 & 0 & 0 & 2 \end{bmatrix}$
[1] TSC	16	$\begin{bmatrix} 0 & 0 & 2 & 1 & 0 & 2 \\ 2 & 1 & 0 & 2 & 2 & 0 \end{bmatrix}$
[1] TSC	32	$\begin{bmatrix} 0 & 0 & 2 & 1 & 3 & 2 & 2 \\ 2 & 1 & 2 & 1 & 3 & 0 & 2 \end{bmatrix}$
[3] BBH	4	$\begin{bmatrix} 2 & 0 & 1 & 3 \\ 2 & 2 & 0 & 1 \end{bmatrix}$
[3] BBH	8	$\begin{bmatrix} 2 & 0 & 2 & 1 & 2 \\ 2 & 1 & 0 & 0 & 2 \end{bmatrix}$
[3] BBH	16	$\begin{bmatrix} 2 & 0 & 1 & 2 & 2 & 0 \\ 1 & 2 & 2 & 0 & 0 & 2 \end{bmatrix}$
[5] CYV	16	$\begin{bmatrix} 1 & 2 & 1 & 2 & 3 & 2 \\ 2 & 0 & 3 & 2 & 2 & 0 \end{bmatrix}$
[4] FVY	16	$\begin{bmatrix} 2 & 0 & 1 & 0 & 0 & 2 \\ 0 & 2 & 2 & 1 & 2 & 2 \end{bmatrix}$

##### C. Minimum trace criterion

Reference [5] derives new design criteria for the case of slow fading channels. It has been shown that for a large product of transmit and receive antenna number ( $>3$ ), the design of codes with maximum coding advantage is determined by the minimum trace of the distance matrix.

$$P(\mathbf{c} \rightarrow \mathbf{e}) = \frac{1}{4} \exp \left( -n_r \frac{E_s}{4N_0} \sum_{i=1}^{n_r} \lambda_i \right)$$

The authors of the same reference: Chen, Yuan and Vucetic present some new codes that maximise the *minimum trace* criterion and are denoted here as CYV codes. However the analytical performance of these codes has not been investigated over the fast fading channel and it will prove useful to include them here.

## V. NUMERICAL RESULTS

The PDU error rate (PER) versus SNR has been adopted here as a suitable measure of performance. The PDU train in the Hiperlan/2 standard contains 54 bytes of data, which is equivalent to 216 4-PSK symbols. It should be noted that such a mode is not actually defined in either 802.11a or Hiperlan/2, however it will prove useful for later comparisons. The aim of this work is to determine the potential of STTCM as applied to OFDM based WLANs. For the simulation studies, the frame is constructed from 4.5 OFDM symbols, where the remaining half a symbol is filled with padding data after the encoding. Using the above configuration, the PDU performance can be compared directly with mode 3 of Hiperlan/2 [6]. In all the following figures, the Channel State Information (CSI) is gleaned by the modem via a sequential transmission of preambles from all Tx elements.

Figure 4 depicts the performance comparison of an FSC-OFDM modem in a configuration using 2 Tx and 1 Rx antenna. The various codes listed in section IV were applied over both measured and synthetic (simulated) channel models. The simulated channels are obtained from independent realizations of ETSI channel A [7]. The synthetic channels have been energy normalized to simulate the best diversity case. Figures 5 and 6 show results for an identical configuration using 2 and 3 Rx antennas respectively.

As expected, the performance over measured channels is worse (compared with synthetic channels) in all cases. The measured case obviously reflects the impact of practical degradations seen in a real channel. The performance gap widens as the number of receive antenna elements increases (and hence as the diversity difference increases). In the synthetic channel the best performing codes are those based on the recently proposed criterion of maximizing the minimum trace of the distance matrix (rather than the minimum determinant). It is interesting to note that using measured channel data this situation is not so clear, and at high SNR quite the opposite conclusion is drawn.

Figures 7 and 8 depict the benefits of applying the codes with a higher number of states. As wideband OFDM systems offer additional diversity across the sub-carriers, higher state order codes can be used to exploit the situation. Most notably, the performance gap between the measured and synthetic cases gets narrower when a higher number of states are used. This demonstrates yet again that the performance loss in the real case is predominantly due to diversity discrepancy in the space domain.

Finally, comparing the results presented here with those obtained using mode 3 of Hiperlan/2 [6], we conclude that FSC-OFDM using 2 Tx and 3 Rx antenna elements can double the system capacity, even over measured channels.

## VI. CONCLUSIONS

In this paper we have investigated the performance of STTCM when applied to OFDM system. Hiperlan/2 and IEEE 802.11a were used as a base system for enhancement. We have developed a signal model for FSC-OFDM. The performance of the proposed system was simulated using state-of-the-art MIMO wideband radio channel profiles. The performance gap between the widely used simulated channel and the measured channel is 4, 5 and 6 dB for 1, 2 and 3 receive antenna elements respectively at  $\text{PER} = 10^{-2}$ .

It has been observed that in the case of measured channel data, the codes of Baro et al. [3] (based on the determinant criterion) performed most consistently at high and low SNR. However, although our results only relate to one measured scenario, the setup is typical enough to conclude the vast potential of STC-OFDM, particularly for use in future WLAN standards.

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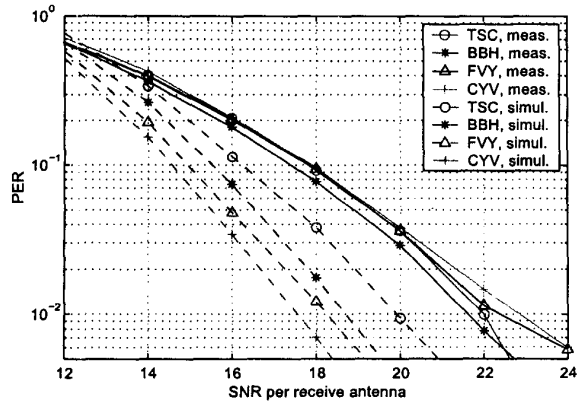


Fig. 4. Performance of different 16-state 4PSK STTCM codes with 1 Rx antenna over measured channels and ETSI channel A.

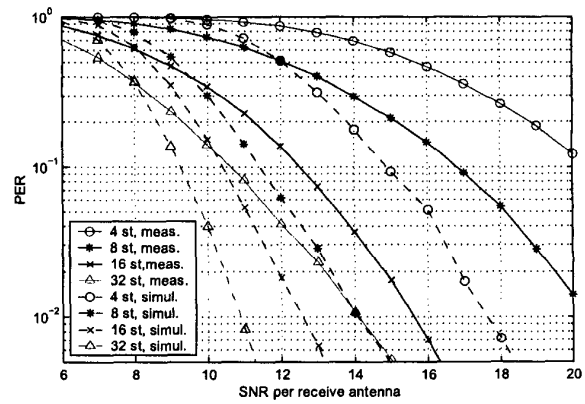


Fig. 7. Performance of 4, 8, 16 and 32 state 4PSK STTCM codes of Tarokh et al with 2 Rx antenna over measured channels and ETSI channel A.

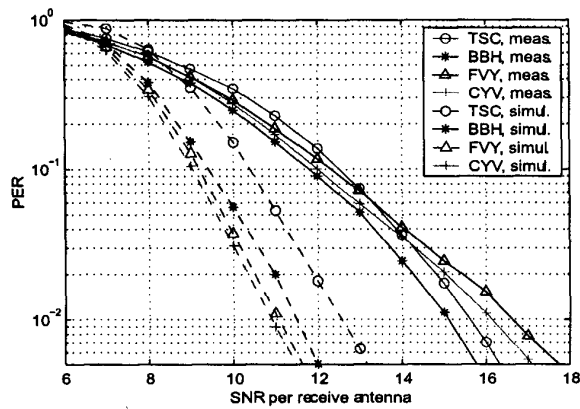


Fig. 5. Performance of different 16-state 4PSK STTCM codes with 2 Rx antenna over measured channels and ETSI channel A.

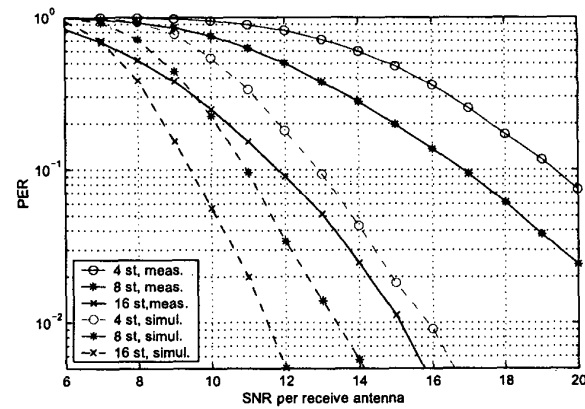


Fig. 8. Performance of 4, 8 and 16 state 4PSK STTCM codes of Baro et al with 2 Rx antenna over measured channels and ETSI channel A.

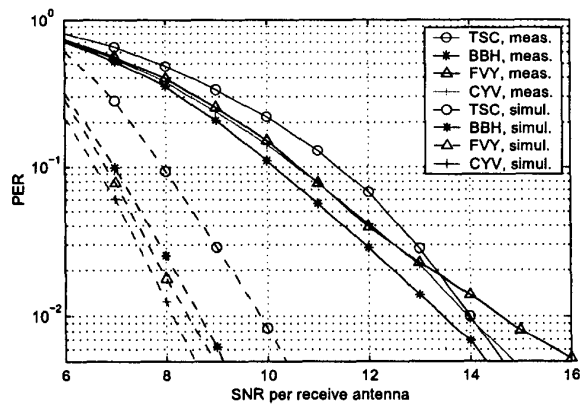


Fig. 6. Performance of different 16-state 4PSK STTCM codes with 3 Rx antenna over measured channels and ETSI channel A.